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# Modulation and Equalisation Considerations for High Performance Radio LANs (HIPERLAN)

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**Abstract: The European Telecommunications Standards Institute (ETSI) is currently defining a European standard for HIgh PERformance Radio LANs (HIPERLAN). To operate as wired LAN replacements, these systems will need to offer data rates as high as 20 Mb/s. To counteract the time dispersive nature of the indoor radio channel, adaptive equalisation is proposed at the receiver.**

**In this paper a number of possible modulation and equalisation techniques are presented and, in particular, the BER performance of quasi-coherent GMSK detection is investigated. Important issues such as frame and symbol synchronisation, frequency offset correction and equaliser implementation are also addressed from a practical standpoint.**

## I - INTRODUCTION

At present, many North American radio LANs operate in the 900 MHz Industrial, Scientific and Medical (ISM) bands. These frequencies are governed by the FCC and are not currently available in Europe. However, many of these radio LANs have been adjusted so that they operate in the more commonly available 2.4 GHz ISM band. At present, these types of product are currently achieving maximum data rates of just 1-2 Mbits/s. To offer a true replacement for current wired-line LANs it is felt that data rates as high as 10-20 Mbits/s will be needed. To address this requirement, ETSI has defined a new Radio Equipment and Systems group (RES-10) whose aim is to draft a European standard for high capacity indoor radio LANs (draft expected early 1995). This standard, known as HIPERLAN, will initially operate in 100 MHz of dedicated spectrum starting at 5.15 GHz. In addition a further 50 MHz of spectrum will be available on a national basis.

In 1992, as part of the European Commission's ESPRIT III program, a project was formed to investigate high data rate transmission techniques for future indoor radio LANs. This project, known as LAURA (Local Area User Radio Access), has worked closely with the standardisation activities carried in ETSI and aims to demonstrate the viability of high capacity radio LANs. This paper focuses on the modulation and equalisation issues relating to both the LAURA and HIPERLAN projects.

Over the years many different techniques have been developed to combat the Intersymbol Interference (ISI) introduced by harsh radio channels. Adaptive equalisation, spread spectrum techniques, Coded Orthogonal Frequency Division Multiplexing (COFDM) and various multicarrier transmission schemes have all been proposed as possible solutions.

As far as the HIPERLAN specification is concerned, the majority of spread spectrum techniques do not appear attractive due to the limited available bandwidth relative to the high transmission rates required. In addition, direct sequence CDMA systems suffer from power control difficulties when ad-hoc network structures are required (as defined in HIPERLAN). For COFDM and multicarrier systems the average signal power is seriously restricted due to CEPT's requirement to restrict the peak envelope power to 1 Watt EIRP (multicarrier systems operate with large peak-to-mean ratios). Moreover, the requirements for the front-end linearity of a multi-carrier based transceiver also impose a number of difficulties which limit the practicality of its implementation. For an equaliser based solution, these linearity and peak power problems can be reduced providing an appropriate modulation scheme is chosen. Equalisation also offers a compact all-digital solution which, when combined with the above arguments, has led to its selection as the preferred technique for both LAURA and HIPERLAN.

This paper presents the issues surrounding the choice of an appropriate modulation and equalisation candidate for HIPERLAN. In particular, issues such as base-band processing and synchronisation are discussed from the LAURA standpoint and simulated Bit Error Rate (BER) results presented for the chosen scheme.

## II - PROPAGATION

To determine the expected indoor values for RMS delay spread at 5.2 GHz, a typical indoor scenario was simplified and used in conjunction with a ray-tracing prediction package [1]. The indoor structure, which was approximately 100m by 100m, was further sub-divided into several smaller rooms. A transmitter was placed in

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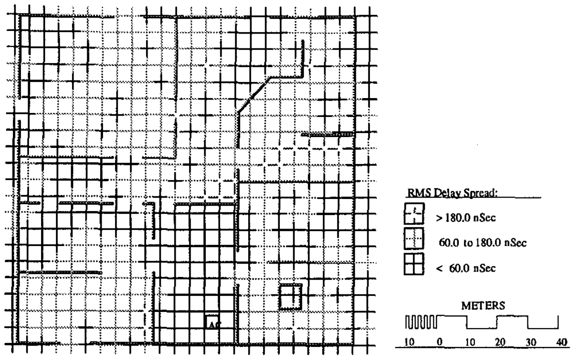



Figure 1: RMS Delay Spread (Indoor)

| Coverage Area | 1% | 5% | 10% | 25% | 50% | 75% | 90% | 95% | 99% |
|---|---|---|---|---|---|---|---|---|---|
| RMS Delay (ns) | 1.5 | 7.4 | 24.5 | 55.1 | 72.9 | 102 | 167 | 192 | 240 |

Table 1: Cumulative RMS Delay Spread

three different locations, figure 1 shows the RMS delay spread map for one of these positions (similar results were also found at the other locations). The results were obtained by calculating the RMS delay spread at each sample point on the indoor grid. The majority of locations experience delay spreads between 50 and 125 ns, the lower values being experienced in line-of-sight locations. However, it is also apparent that small areas do exist with RMS delay spread values greater than 150 ns (although many of these are outside the coverage area of the node). To understand the variation of delay spread over the entire building, table 1 shows the cumulative results. An equaliser capable of supporting up to 150 ns should be able to handle almost 90% of locations. However, even if only 100 ns of RMS delay could be tolerated, 75% of locations would be supported and these would include almost all line-of-sights conditions.

The above ray-tracing analysis was also used to generate the channel models used in the later simulations. The software determines all the significant rays travelling from transmitter and receiver. The resulting complex channel impulse response is then incorporated into the system and used to determine the tap weightings in the channel model.

## III - MODULATION

The LAURA and HIPERLAN air interface techniques use adaptive equalisation to overcome the harmful effects of ISI. The choice of modulation is governed by a number of competing factors, the most important being: transceiver linearity, spectral efficiency, interference tolerance and equaliser complexity [2][3].

Linear modulation schemes such as $\pi/4$-QPSK and OQPSK achieve high levels of bandwidth efficiency through the use of root-raised cosine filtering, however their need for linear amplification effectively rules them out of contention on the grounds of required amplifier linearity. Amplifier linearity is required to prevent excessive Inter-Modulation Distortion (IMD) from spreading the spectrum outside of the specified frequency mask. In ad-hoc radio networks where stations are scattered randomly in the radio coverage area, it often happens that a transmitter sending unwanted signals is much closer to a reference receiver than the one transmitting wanted signals (the so called "near-far" scenario) [4]. In such cases, the IMD from the unwanted transmitter will fall into adjacent channels and, if sufficiently high, will cause the unwanted blocking of neighbouring channels. Although suitable performance may be obtained by backing off currently available amplifiers, the power efficiency of such an approach is clearly not compatible with battery power PCMCIA applications. Given the cost and complexity of a linearised High Power Amplifier (HPA) at 5.2 GHz, it was concluded that variable envelope modulation schemes like root-raised cosine QPSK could not be supported [2][3].

Constant envelope modulation schemes such as Continuous Phase Frequency Shift Keying (CPFSK) were then considered (see figure 2), namely GMSK and m-ary CPFSK (m=4,8). Higher level CPFSK schemes are clearly desirable from the point of view of lowering both transceiver and equaliser complexity (a lower symbol rate produces shorter equaliser spans and lowers the required sampling rate), however the non-linearity introduced by the bandlimiting filters can seriously degrade the performance of symbol-by-symbol based DFE detection. Figure 3 shows graphically the impact of frequency shaping on two types of CPFSK transmission. Figure 3(ii) illustrates that as a result of the frequency shaping LPF,

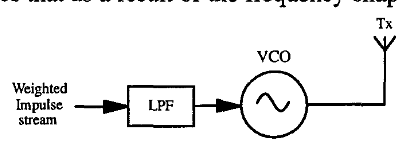
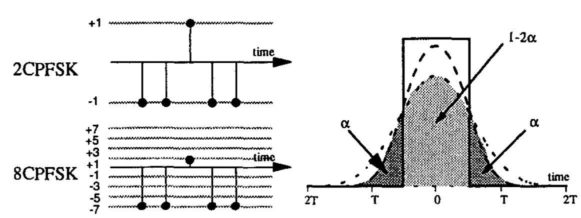



Figure 2: CPFSK Transmitter




(i) Impulse Response (ii) Premodulation Filter

Figure 3: Premodulation Filters for CPFSK

some of the wanted symbol spills over into neighbouring time slots (ISI). Since the shaping is in the frequency domain the resulting distortion is non-linear as far as the received phase is concerned. The degree of frequency ISI can be determined by the area in adjacent symbol slots, $\alpha$. As $\alpha$ increases, the resulting eye diagram in the receiver begins to close, as the degree of filtering is increased there comes a point where symbol by symbol detection is no longer possible. The worst case symbol streams are shown below with the resulting weightings for the wanted symbol. For the case of MSK, the wanted symbol has a weight of $1\text{-}2\alpha$ and, in the worst case, the neighbouring symbols each subtract a weight of $\alpha$ yielding a final result of $1\text{-}4\alpha$

$$\text{GMSK} \quad -1,+1,-1 \rightarrow 1-4\alpha \quad \text{Max } \alpha = 0.25 \qquad (1)$$

$$\text{4CPFSK} \quad -3,+1,-3 \rightarrow 1-8\alpha \quad \text{Max } \alpha = 0.125 \qquad (2)$$

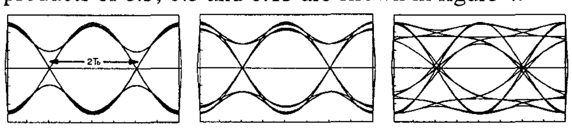
$$\text{8CPFSK} \quad -7,+1,-7 \rightarrow 1-16\alpha \quad \text{Max } \alpha = 0.063 \qquad (3)$$

For symbol by symbol detection, the resulting weighting must remain greater then zero if the received eye diagram is to remain open. Hence, for 2, 4 and 8CPFSK the maximum value of $\alpha$ will be 0.25, 0.125 and 0.063 respectively. In practice, to achieve reasonable performance, the value of $\alpha$ should be well within this maximum value. Table 2 shows the value of $\alpha$ for various types of premodulation filter.

| Filter Type | $\alpha$ | Filter Type | $\alpha$ |
|---|---|---|---|
| Gaussian BT=0.2 | 0.228 | Gaussian BT=1.0 | 0.054 |
| Gaussian BT=0.3 | 0.107 | Gaussian BT=1.5 | 0.037 |
| Gaussian BT=0.5 | 0.071 | 3RC | 0.194 |

Table 2: ISI weighting for Various Filters

For GMSK, a Gaussian filter with a BT product of 0.5 can be seen to result in a small value of $\alpha$, hence the resulting non-linear ISI and eye-closure will be small. As the BT value is lowered below 0.3 the value for $\alpha$ increases rapidly and detection becomes more complex. The resulting I and Q eye diagrams for GMSK with BT products of 0.5, 0.3 and 0.15 are shown in figure 4.




(i) BT=0.5 (ii) BT=0.3 (iii) BT=0.15

Figure 4: Eye diagrams for GMSK

From the above it can be seen that for BT values greater than 0.3, the resulting eye diagram is not dissimilar to that of OQPSK. For higher levels of CPFSK, the degree of tolerable non-linear ISI is reduced, hence if the bandwidth is to be contained, some form of sequence estimation in the receiver may become necessary. Unfortunately, the complexity of implementing sequence estimation is generally considered too great at 10-20 Mbits/s and thus a 2-CPFSK solution was chosen.

GMSK was found to offer a good performance, even when frequency pulse shaping was present (BT=0.5). Since the received I and Q eye-diagrams closely approximate to OQPSK, LAURA makes use of data precoding in the *transmitter* to allow GMSK to be received as OQPSK, see figure 5 below.

| | | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| Data | +1 | +1 | -1 | -1 | +1 | -1 | +1 | -1 | -1 | -1 | +1 | +1 | -1 | +1 |
| $\Omega t$ | 90 | 180 | 270 | 0 | 90 | 180 | 270 | 0 | 90 | 180 | 270 | 0 | 90 | 180 |
| Toggle $d_n$ | +1 | -1 | +1 | -1 | +1 | +1 | -1 | -1 | -1 | +1 | -1 | +1 | -1 | -1 |
| XNOR[$d_n$,$d_{n-1}$] | +1 | -1 | -1 | -1 | -1 | +1 | -1 | +1 | +1 | -1 | -1 | -1 | -1 | +1 |
| Phase ($\theta_n$) | 90 | 0 | 270 | 180 | 90 | 180 | 90 | 180 | 270 | 180 | 90 | 0 | 270 | 0 |
| cos $\theta_n$ | 0 | +1 | 0 | -1 | 0 | -1 | 0 | -1 | 0 | -1 | 0 | +1 | 0 | +1 |
| sin $\theta_n$ | +1 | 0 | -1 | 0 | +1 | 0 | +1 | 0 | -1 | 0 | +1 | 0 | -1 | 0 |
| $b_n$ | +1 | +1 | -1 | -1 | +1 | -1 | +1 | -1 | -1 | -1 | +1 | +1 | -1 | +1 |

Figure 5: GMSK Pre-Coding Table

In the transmitter, two bits in every four are toggled and the resulting bit stream, $d_k$, passed through an XNOR function. The resulting GMSK transmission can now be received as QPSK with the sine and cosine of the received phase directly corresponding to the binary data. The removal of the differential aspects of GMSK is important if Decision Feedback Equalisation is required without the need to remodulate the incoming data.

The resulting constant envelope GMSK/OQPSK solution produces a BER performance that is very close to that of coherent QPSK, in AWGN a degradation of less than 1 dB is observed.

## IV - EQUALISATION

Various equalisation strategies can be applied at the receiver, in order of complexity the following three schemes were considered: Linear Transversal Equalisation (LTE), Decision Feedback Equalisation (DFE) and Viterbi Sequence Estimation (VE). Of these schemes, DFE equalisation was chosen as a compromise between performance and complexity. T/2 spacing is used in the feedforward filter to reduce timing sensitivity (note: in this case T refers to the GMSK bit period).

To obtain optimal performance and minimise computational complexity, the number of feedforward and feedback taps must be chosen carefully. In general this depends on the power delay profile and the maximum

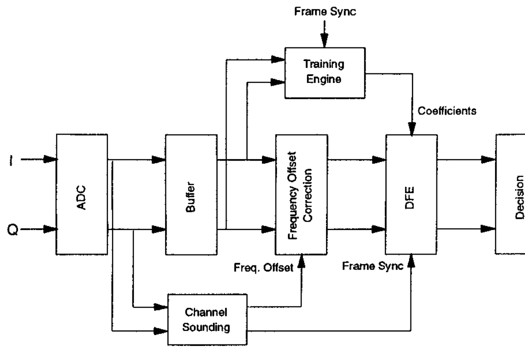



Figure 6: Off-Line Equalisation

| | | | | | | | |
|---|---|---|---|---|---|---|---|
| Start Sequence | Training Sequence | Data | Sounding Sequence | Data | Sounding Sequence | Data | Tail Sequence |

Figure 7: Packet Structure

delay spread. A channel impulse response can be broken down into pre-echoes, a main component and post echoes. The equaliser should be configured so that all the post echoes are eliminated by the feedback filter and the pre-echoes by the feedforward filter. Channel sounding is performed during the initial synchronisation and this allows the number of feedback and feedforward filter taps to be determined (more realistically, these taps would be fixed at their maximum span length).

Real time equalisation was considered impractical, even with the simplest training algorithms [5]. Hence, "Off-line" equalisation has been implemented in LAURA. A block diagram illustrating this approach is shown in figure 6 [6]. Figure 7 shows the packet structure used with this form of equaliser. The coefficient update algorithm is implemented in LAURA using fast DSP. The feedforward filter is implemented with a standard programmable FIR filter while the feedback section uses look-up tables directly computed from the training engine coefficients (for HIPERLAN these tasks would be integrated into a custom ASIC design).

Synchronisation signals (see later section) allow the training sequence to be captured and read into the DSP hardware, while the DSP is performing the coefficient update, the incoming signals are continuously buffered. For the LAURA design, the buffer is larger enough to accommodate all the incoming signals during training (for HIPERLAN, this buffer would be small and certainly no larger than the smallest packet size supported by the protocol). After training, the resulting co-efficients are down-loaded into the feedforward and feedback filters.

Since "Off-line" training is used it is no longer valid to measure the algorithm complexity using operations per iteration. Instead, the total number of operations (i.e. the operations per iteration multiplied by number of iterations) should be used for algorithm complexity measurements. Using these measures, Kalman algorithms become comparable to LMS techniques. There are several forms of Kalman algorithm, the square-root solution being preferred to the "fast" algorithm due to its lower sensitivity to rounding error [7]. This algorithm requires $1.5N^2 + 5.5N$ complex multiplications and N complex divisions, where N represents the number of coefficients in the equaliser. For an 11 tap filter, 253 complex operations are required per iteration and, assuming 48 iterations, a total of 12,144 complex operations are require to train (for a 17 tap filter this value increases to 26,112). In LAURA, four parallel TMS320C40 processors are being used to train each packet in approximately 3 ms. Custom equaliser design should easily reduce this delay to a more acceptable value. If the smallest packet size contains just 512 bits and a data rate of 20 Mbits/s is assumed, a single packet will last for 25.6 μs. Hence, for 11 and 17 tap equalisers, 454 and 1020 million complex instructions per second are required to train in the smallest packet size.

## V - SYNCHRONISATION

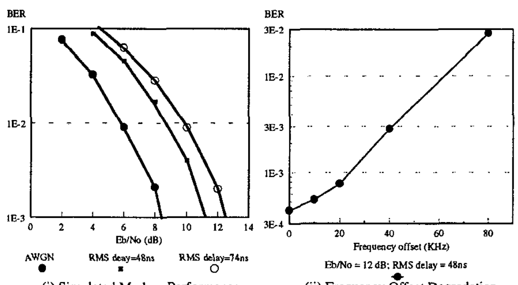
In general, knowledge of the exact carrier and symbol phase is required in the receiver. However, since T/2 spaced equalisation is being used, the initial training algorithm will handle both phase offsets and timing errors. However, carrier references need to be stable enough to ensure that the maximum phase rotation is small during a frame. In practise this would result in very expensive synthesisers and so LAURA makes use of regular phase estimation and correction sequences.

Symbol timing is achieved with a free running clock (due to its low frequency), however the carrier frequency offset and frame synchronisation need to be considered carefully. Due to limited carrier frequency stability in both the Tx and Rx oscillators, there is always a residual frequency offset in the received base-band signal. If the frequency offset is large, the DFE will not work efficiently due to rapid phase rotation. Good synchronisation between the incoming training sequence and the local reference sequence is also vital for effective training. Synchronisation error decorrelates the two signals and may result in degraded performance or even the total collapse of the equaliser.

LAURA uses a wideband channel sounding sequence to achieve both frame synchronisation and frequency error compensation. The principle of this technique is well known, by transmitting a specially-designed sequence, the overall channel characteristics in I and Q can by generated by a complex correlation process at the receiver. The amplitude peak of the correlation can be used for frame synchronisation. Frequency offset compensation is achieved based on analysis of successive correlation phases. For GMSK, a binary sequence has

| Modulation | Pre-coded GMSK with BT = 0.5 |
|---|---|
| Bit Rate (Gross) | 20 Mbits/s |
| Training Sequence | 48 bits |
| Filter Span | 6 feedforward + 5 Feedback |
| Training Algorithm | Square Root RLS (Kalman) |
| Channel RMS Delay | 48 ns and 74 ns |
| Synchronisation Sequence | 26 bits repeated every 326 bits |

Table 3: Simulation Parameters




(i) Simulated Modem Performance (ii) Frequency Offset Degradation

Figure 8: Preliminary DFE Performance

been designed so that the modulated I and Q channels produce two sequences that have good auto-correlation and low cross-correlation [8]. At the receiver, by correlating the incoming I and Q signals with the local sequences, channel characteristics in I and Q can be generated. Hence, by repeating the sounding sequence at an appropriate frequency, channel phase rotation (and therefore frequency offset) can be adequately monitored and removed using a Numerically Controlled Oscillator.

## V I- SIMULATION RESULTS

To demonstrate the performance of the LAURA proposal, a software simulation with the specifications shown in table 3 was implemented (Note: larger equaliser spans are required for non-line-of-sight scenarios, LAURA caters for equaliser spans up to 12 feedforward and 5 feedback taps) .

The channel impulse responses used in these simulations were generated by ray-tracing the indoor office-like scenario shown in figure 1. Figure 8(i) shows the performance of the equaliser in two line-of-sight locations (Note: these simulations incorporated the proposed phase, frequency, symbol and frame synchronisation algorithms and did not assume any prior information from the transmitter). The relatively poor performance at 74 ns RMS delay spread arose due to the short length of the feedforward section which spans just 132 ns. Simulations also investigated the performance of equalisation in the presence of frequency offset. It is clear that using a training sequence with a frequency offset leads to degraded performance. Figure 8(ii) shows the impact of frequency offset on the DFE training process. This graph shows that frequency offsets up to 20 kHz can be tolerated without correction. If higher frequency offsets are present then correction will have to take place before training and this will add extra delay and complexity.

## VII- CONCLUSIONS

This paper has summarised the techniques used in the LAURA modem. Pre-coded GMSK was selected as the best compromise between linearity, spectral efficiency and equaliser complexity / performance.

The performance of the proposed packet based system has been simulated and its BER determined for two typical indoor channels. HIPERLAN is expected to operate with a data rate of around 20Mb/s. To conserve bandwidth, higher level modulation schemes such as 4 and 8CPFSK were considered. However, as shown in this paper, such schemes suffer badly from non-linear ISI introduced in the frequency shaping filters and, as a result, their performance is either inferior relative to GMSK, or requires highly complex sequence-based equalisers. In conclusion, precoded GMSK offers the most viable solution for high capacity indoor data transmission.

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